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SYSTEM ARCHITECTURE STUDY OF AN ORBITAL GPS USER TERMINAL

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TABLE OF CONTENTS

TABLE OF CONTENTS

FIGURES

CHAPTER 1

INTRODUCTION

This report presents work accomplished under contract NAS5-25819. The primary objectives of this study are the following:

- Define the generic RF and applications processing requirements for a GPS orbital navigator;

- Determine a reasonable line of demarcation between dedicated analogue hardware, and software/processor implementation, maximizing the latter;

- Describe a modular approach to R/PA design which permits several varieties of receiver to be constructed from basic components.

It is a basic conclusion of this investigation that software signal processing of the output of the baseband correlator (Figure 3-3) is the best choice of transition from analogue to digital signal processing. High-performance sets requiring multiple channels are developed from the generic design of Figure 3-3 by replicating the RF processing segment, and modifying the applications software to provide enhanced state propagation and estimation.

In this report, Chapter 2 discusses GPS RF signal processing requirements; Chapter 3 describes applications requirements; and Chapter 4 presents the partitioning analysis.

CHAPTER 2

RECEIVER PROCESSOR SIGNAL PROCESSING

2.1 Signal Processing

This section will review the GPS signal structure, and outline the signal processing necessary to recover the relevant information contained in the GPS signal. The following subsections discuss GPS signal processing requirements in greater detail. Section 2.1.1, "Signal Collection," is a discussion of the radio frequency processing required to obtain a usable signal. Section 2.1.2, "Detection and Tracking," describes the generation and maintenance of the local replica of the GPS signal. Finally, Section 2.1.3, "Data Demodulation," outlines the processing required to recover the navigation data message as well as range and doppler (range rate) information.

The GPS signal consists of two separate L-band signals called Link 1 (L1); and Link 2 (L2). The L1 signal has a nominal carrier frequency of 1575.42 MHz, while the L2 signal has a nominal carrier frequency of 1227.6 MHz. The L1 signal consists of two components in phase quadrature. The first is bi-phase modulated with the modulo-2 sum of a 50-bit per second binary data message and a 10.23 MHz binary pseudo-random number (PN) sequence known as the precision code for P code. The second component of the L1 signal is similarly modulated with the data message and a 1.023 MHz PN code known as the Clear/Acquisition code or C/A code. The L2 signal is simply modulated with the data message and P code. See Figure 2-1.

$$S(t) = S_{L1}(t) + S_{L2}(t)$$
$$S_{L1}(t) = \sqrt{2S_{p1}}\, d(t)p(t)\cos w_{L1}t + \sqrt{2S_{c1}}\, d(t)c(t)\sin w_{L1}t$$
$$S_{L2}(t) = \sqrt{2S_{p2}}\, d(t)p(t)\cos w_{L2}t$$

where:

$S(t)$ = GPS satellite signal

$S_{L1}(t)$ = Link 1 signal (L1)

$S_{L2}(t)$ = Link 2 signal (L2)

S_{p1} = Power in P-code (inphase) component of L1 signal

S_{c1} = Power in C/A-code (quadrature) component of L1 signal

S_2 = Power in L2 signal

$d(t)$ = Data modulation (50 BPS)

$p(t)$ = This satellite's Precision code (10.23 MHz)

$c(t)$ = This satellite's Clear/Acquisition code (1.023 MHz)

W_{L1} = $2\pi\ 154\ f_o$

W_{L2} = $2\pi\ 120\ f_o$

f_o = 10.23 MHz

NOTES: 1) $d(t)$, $p(t)$, and $c(t)$ take on a value of ±1

2) Modulation on L2 may be changed from P to C/A by ground command

FIGURE 2-1 GPS NAVIGATION SIGNAL STRUCTURE

All frequencies in the GPS signal structure, carriers as well as modulation rates, are coherently derived from a single frequency, f_o, supplied by the GPS satellite frqeuency standard. This frequency is nominally[1] 10.23 MHz. The L1 carrier is 154 f_o, L2 is 120 f_o. The transition rate (or "chipping rate") of the P code is f_o, and for the C/A code the chipping rate is $f_o/10$, etc.

The ability of a GPS receiver to detect the GPS signal, to differentiate among several GPS signals, to determine the time of transmission of a particular signal, and to demodulate the navigation data message is inherent in the correlation properties of the PN modulation on the signal. Figure 2-2 outlines the general correlation properties for both P and C/A codes, and [SPILKER 78] discusses these properties in detail. Any GPS receiver will, of necessity, generate a PN code replica and correlate it with the received signal as part of its signal processing.

A generic GPS receiver will perform the following signal processing. First, one of several possible antennas is selected which has the satellite in view. The signal received by this antenna is fed first to an L-band pre-amplifier and then to a frequency down-conversion stage. One or more stages of band pass filtering usually occurs in the amplification and frequency down-conversion process--but the information inherent in the signal, that is, carrier phase and modulation, is always preserved. This is what will be referred to as "Signal Collection".

The next step, "Detection and Tracking," consists of generating a replica of the PN code to be tracked, and correlating it with the received signal. The output of the correlation process is used to maintain the signal replica, usually by use of a delay locked loop. Simultaneously, the residual carrier is removed, either by use of a tracking loop or by envelope (or power) detection, or both. Pseudo-range is then available as a result of tracking the PN code (other P or C/A), and pseudo-delta range is available as a result of coherent carrier tracking. The data modulation is available whenever both the PN code and the carrier phase are being tracked.

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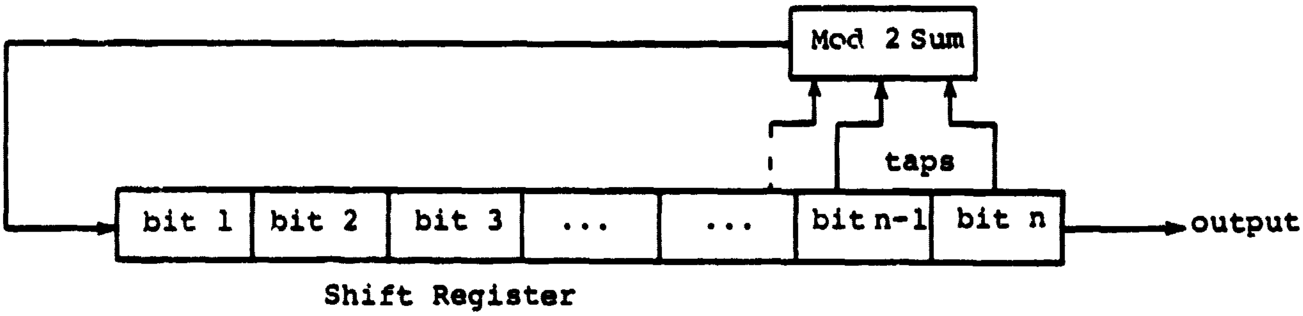
[1] There is a relativistic correction of the transmitted frequency at the satellite. f_o is nominally 10.23 MHz when measured on the surface of the earth.

- Both Precision and Clear/Acquisition codes are variants of the Linear Feedback Shift Register sequence (LFSR)



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Shift operation generates next bit (and shift register state) in sequence.

Correct choice of taps and initial shift register state will generate all 2^n states but one - the zero state is degenerate.

Period of sequence (n_p) is then 2^n-1

- Autocorrelation of a LFSR sequence:

s(t) - LFSR sequence with shift time t_c (one chip)
= ±1

$$\text{Autocorrelation} - \phi(\tau) = \frac{1}{n_p t_c} \int_0^{n_p t_c} S(t)S(t+\tau)dt$$

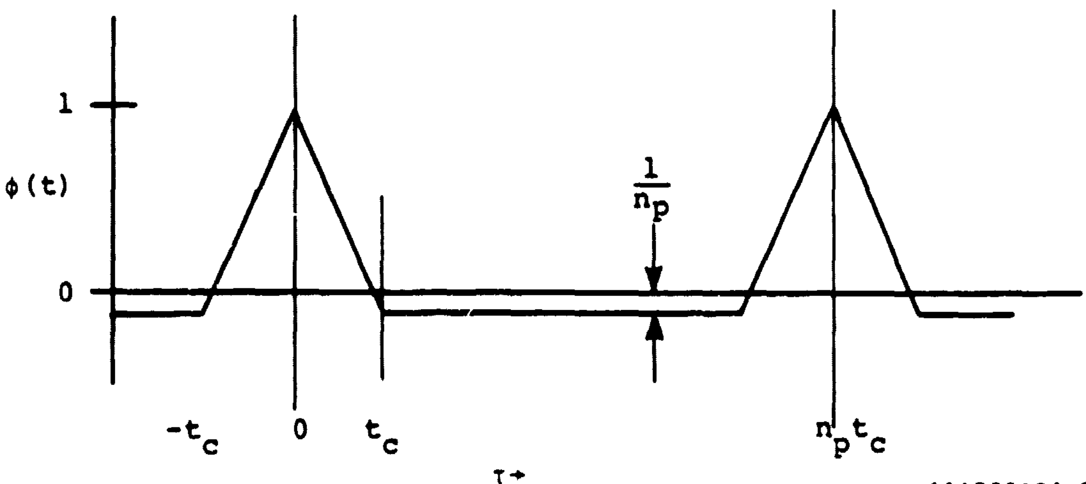

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FIGURE 2-2 GPS PN CODE CORRELATION PROPERTIES

2.1.1 Signal Collection

This section will discuss the generic functions of a GPS receiver required to obtain a usable signal. The most important of these functions is the antenna; a GPS receiver cannot do without one. The remainder of the functions discussed in this section are secondary in nature--they are not strictly necessary for the conceptual realization of a GPS receiver. In fact, these functions have one thing in common: they do not change the information contained in the GPS signal. They are important from the standpoint of a practical realization of a receiver, however.

Figure 2-3 is an overview of the functions discussed in this section. The antenna is generally followed by a preamplifier. The function of the preamplifier is to provide signal gain to overcome losses due to cabling runs. A subsidiary function is to filter wide-band noise. After the antenna and preamplifier, an antenna selection function is required in the case of multiple antennas. Shown next is calibration signal injection. The calibration signal is a duplicate of a GPS signal but uses a PN code known not to be that of any existing satellite. The injection of this signal allows the receiver to determine ambient noise levels at the antenna, perform automatic gain control (AGC) initialization, calibrate internal signal delays, etc. The last function shown is that of frequency conversion. This function translates either the L1 or L2 signal to a common intermediate frequency. This determines whether a subsequent stage tracks L1 or L2.

2.1.1.1 Antenna and Antenna Selection

The GPS signal is a circularly polarized L-band signal. Potential emitters are widely separated in space, generally on a hemisphere centered on the receiver (at least for an earth-bound receiver). This defines the most commonly used GPS anenna system--a single L-band antenna, suitable for circularly polarized signals with a hemispherical gain pattern.

Requirements for greater performance, and consideration such as multipath and jammer rejection, antenna shadowing, etc., have led to the use of more sophisticated systems. In the case of an orbital user in a non-hostile environment, it is expected that principal considerations would be: 1) a priori initiation is an antenna placement on the satellite, and 2) attitude stabilization of the satellite (space stable, earth level, solar pointing, etc.). A typical antenna system might then consist of more than one antenna placed to give roughly spherical coverage. This, then,

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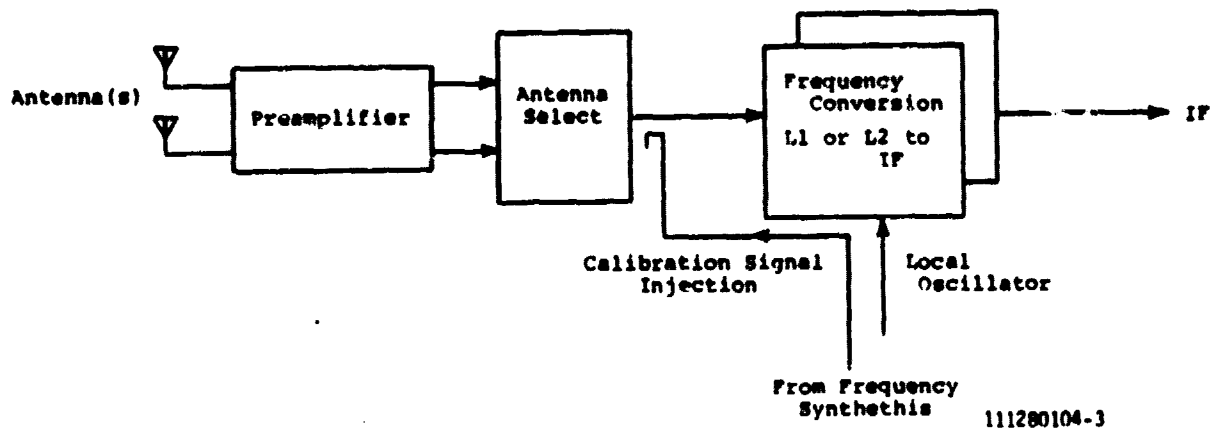


FIGURE 2-3 SIGNAL COLLECTION

requires the ability to select the antenna with the desired GPS satellite in view. In addition, the GPS navigator must have some knowledge of vehicle attitude.

2.1.1.2 Preamplifier

As mentioned above, the objective of the preamplifier stage is immediate gain to prevent the signal from being "lost" in the thermal noise it will encounter in subsequent stages. Other subsidiary functions are bandpass filtering to reduce broadband noise. The signal may also be amplitude limited at this stage. Figure 2-4 is a block diagram of a typical preamplifier stage. Notice that it actually consists of two bandpass filters and anplifiers--one for L1, another for L2. The pass band of these filters must be wide enough to pass the PN modulation on the signal and to accommodate any doppler shift.

This preamplifier would be typical for a receiver which tracks the precision code. This is implied by its ability to handle both L1 and L2. (Typically, a C/A code only receiver cannot track L2, as it carries only P-code.) One would assume, then, that bandwidth of the filters shown is slightly larger than 20.46 MHz to pass both P-code and signal doppler. A C/A only set would need only the L1 arm, and its bandwidth would ideally be about 2.046 MHz. Clearly, a preamplifier designed for a P-code tracking receiver would not necessarily be the "optimal" one for a C/A only receiver, unless its design were complicated by having selectable bandwidths. (This discussion ignores the practicality of constructing a stable bandpass filter centered at 1.5 GHz with a bandwidth of only 2 MHz. If this is in fact difficult or expensive, one would simply use as narrow a bandwidth as practicable. The argument applies more strongly in the discussions of IF stages that follow.)

2.1.1.3 Frequency Down-Conversion

Virtually any practical GPS receiver will have one or more stages of frequency down-conversion in order to facilitate further processing of the signal. A typical stage is shown in Figure 2-5.

This stage is essentially a standard intermediate frequency (IF) stage, in which all local oscillator (LO) frequencies are synchronously derived from the receiver's common clock. A synchronous local oscillator is required to preserve the phase information in the input signal for later stages.

Another function shown in Figure 2-5 is that of automatic gain control. Both local and external AGC

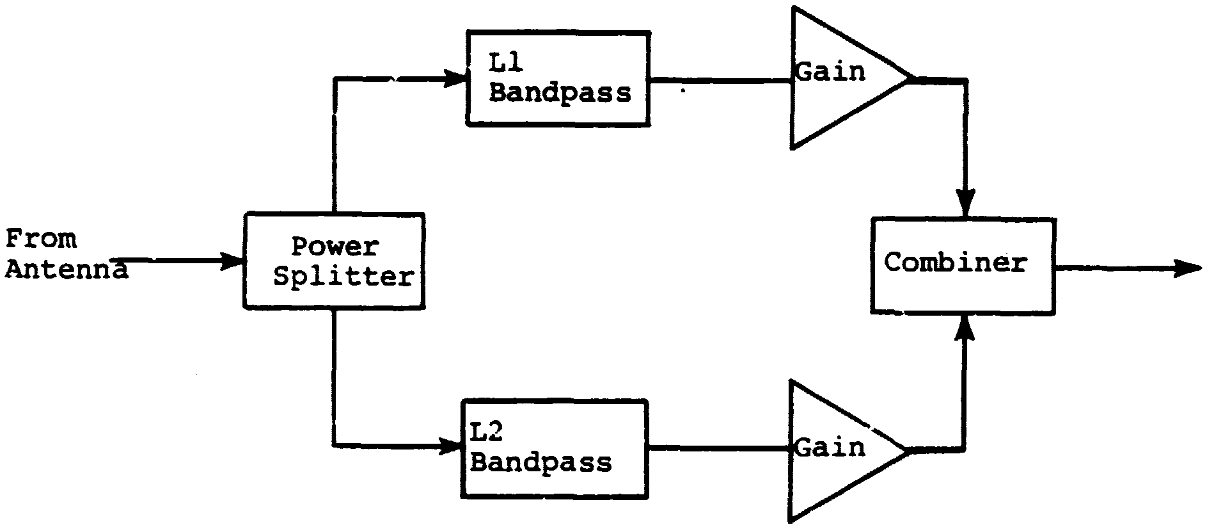


FIGURE 2-4 PREAMPLIFIER

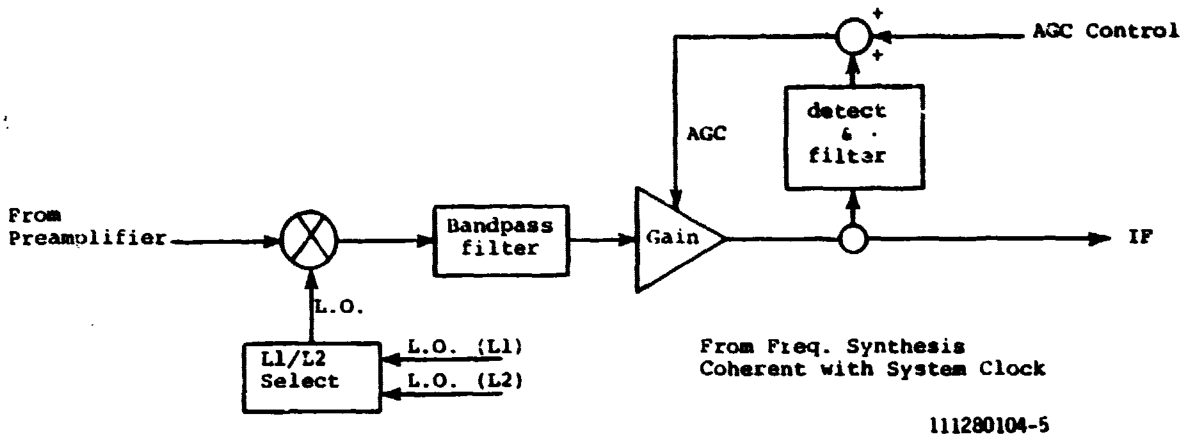


FIGURE 2-5 FREQUENCY DOWN-CONVERSION

controls are shown. AGC will be discussed in a later section, as it plays an important role in the operation of tracking loops and acquisition logic.

In addition to frequency translation and AGC, the third function shown is that of L1/L2 selection. By selecting the appropriate L.O. frequency, either L1 or L2 is translated to the single IF frequency.

The bandwidth of this stage, as in the case of the preamplifier stage, must be at least as wide as the code to be tracked plus any doppler on the incoming signal. To obtain the best performance in the presence of noise, it is desirable to obtain this limit. Thus, if the set is to track C/A only, a 2 MHz bandwidth is appropriate. If the set is to track P-code, a 20 MHz bandwidth is appropriate.

Some current receiver designs use a doppler shifted L.O., that is, the local oscillator frequency is derived from the tracking loops estimate of the incoming carrier frequency. This reduces the bandwidth requirement to only that of the code modulation, as the residual doppler is "removed" at the mixer stage. This bandwidth reduction can improve performance in this presence of noise. This increases the complexity of the frequency synthesis performed by the receiver, and the reduction in noise bandwidth is relatively small, that is, 100 kHz for typical orbital navigator doppler shifts, as compared to 2.045 MHz for C/A code and 20.46 MHz for P-code.

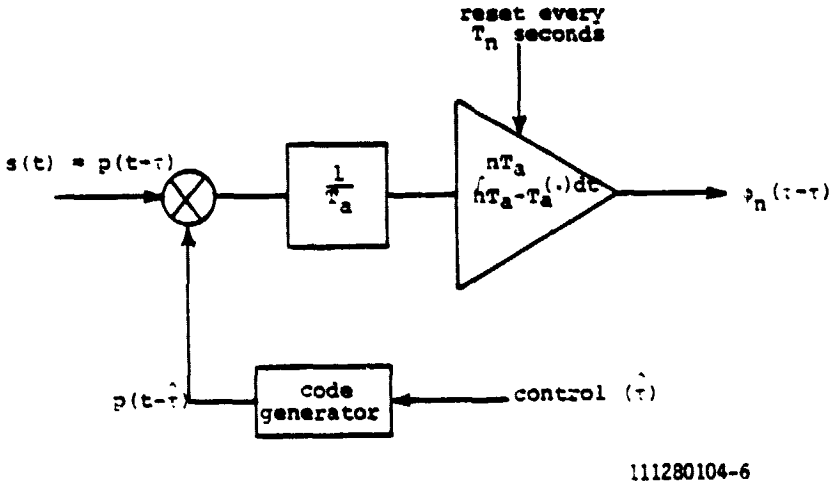
It should also be noted that there may be more than one stage of frequency conversion in a realizable GPS receiver design.

2.1.2 Signal Detection and Tracking

The preceding sections have discussed signal processing that essentially leaves the information content of the received signal unchanged--that is, amplification, frequency translation, and gain control. This section will discuss the signal processing required to detect or demodulate that information.

The GPS navigation signal utilizes a suppressed carrier, spread spectrum modulation technique. (This refers to the "frequency spreading" properties of the PN code.) Signal detection for such a signal can always be characterized as follows: 1) generate a local replica of the incoming signal, and 2) correlate the incoming signal with the local replica. From the correlation properties of PN codes discussed previously, one sees that when the local replica epoch matches that of the incoming signal, the result of correlation will be a maximum.

There are essentially two distinct methods of performing the correlation mentioned above. The method that has been used in GPS receivers to date is that of "active correlation". This involves generation of the code replica in sequential fashion, multiplying this replica with the incoming signal and integrating the product. This technique implies the ability to control the code generator's epoch to bring it into line with that of the incoming signal. Also, note that a true correlation would imply an infinite time average of the product, or for a periodic code sequence, integration over one period. Figure 2-6 depicts a practical active correlator. Here, an integrator is used to perform a finite time average in order to approximate a true correlator. One output is produced every T_a seconds. Other practical implementations exist. For example, the finite time averager could be replaced with a first order low-pass filter. Such an implementation would produce a continuous output that would be an exponentially weighted time average.



$p(\cdot)$ = code sequence, taking on values ± 1

t = user time

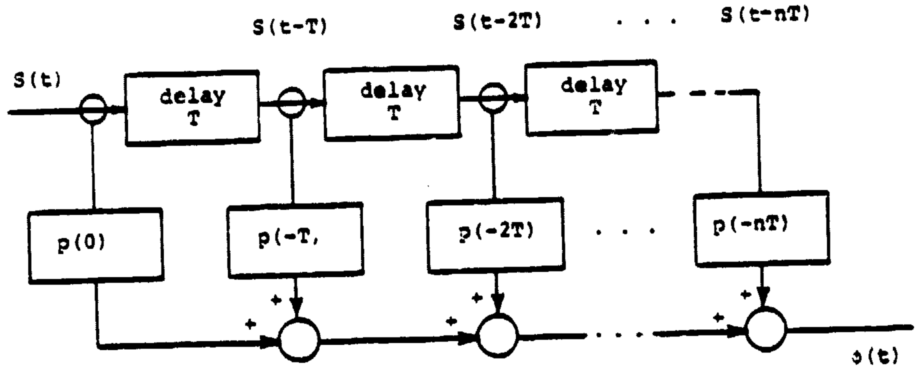
τ = code delay with respect to user time

$\hat{\tau}$ = local estimate of delay

T_a = averaging time

FIGURE 2-6a ACTIVE CORRELATOR

The second technique to be considered is that of "passive correlation". This technique uses what is sometimes called a "matched filter," or in digital signal processing, a "finite impulse response filter," so called because its impulse response is identically equal to zero after a finite period of time. In the passive correlator shown in Figure 2-6b the output, $\phi(\underline{\alpha})$, is equal to n (the number of stages, or order of the filter) when the input is



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$$\phi(t) = \sum_{k=0}^{k=n} S(t-kT)\,p(-kT)$$

FIGURE 2-6b PASSIVE CORRELATOR

at the start epoch of the code sequence, p(o). At other times, the output would be approximately zero. (In the figure, the delay interval, T, is the duration of one bit of the sequence, or an integral fraction of one bit.)

As an example, a passive correlator for the C/A code might consist of 1023 stages, one for each bit of the C/A code sequence. The bit sequence of a particular C/A code to be tracked would comprise the gains shown in Figure 2-6b. The output of such a filter would be a series of pulses, spaced one millisecond apart, occurring when the input signal starts the C/A sequence. The time at which these pulses occur relative to user time would constitute the delay measurement. The P-code would present problems with this method, as it repeats itself only once per week. A solution to this problem would consist of periodically changing the filter's gains to correspond to a part of the P-code sequence to occur in the future, for example, at one millisecond intervals. Alternatively, the gains themselves could be held in a shift register or delay line, and be fed from a clocked code generator, as in the active filter.

A practical passive correlator for GPS signal detection would require a large number of stages to achieve reasonable integration times. For example, to achieve the same performance as an active correlator with a one millisecond averaging time, 1023 stages would be needed for C/A code, and 10,230 stages for P-code.

One advantage of the passive correlation technique is the lack of a requirement for acquisition logic. Where an active correlator must search for the signal if its output is initially near zero, a passive correlator will "find" the code within one code sequence. For example, an active correlator search for the correct C/A code epoch may require spending a millisecond or so, searching each of 1023 "code bins," whereas a passive correlator will produce usable output within one millisecond. This drawback of active correlation can be ameliorated in two ways: 1) good a priori knowledge of time and signal delay. This will reduce acquisition time, but in general will not eliminate the requirement for some type of search strategy; and 2) combining active and passive techniques, as in the use of multiple correlators performing parallel searches. Again, this will reduce acquisition time, but it will probably increase hardware requirements.

The advent of several new technologies such as Charge Coupled Devices (CCDs), Surface Acoustic Wave devices (SAWs), and Very Large Scale Integration (VLSI) fabrication techniques for digital matched filters, have recently made relatively large passive correlators a practicality. The GPS signal structure still places stringent requirements on the passive correlator, particularly a high chipping rate (1 or 10 MHz) combined with long code sequences and low received power. While passive correlators may find uses in GPS receiver design such as acquisition aiding, it appears unlikely that they will replace active correlation techniques in the near term. In particular, the high doppler shifts that would be encountered by an orbital navigator cause difficulties for the matched filter technique (see CAHN73, p. 5-46). The remainder of this section will therefore assume that an active correlator will be used by the receiver.

Figure 2-7 depicts an active correlator and phase detector as might be used in a practical receiver. The incoming signal is first mixed with a doppler shifted local oscillator, then with the PN code that is to be demodulated (either C/A or P for the emitter to be tracked). This is followed by a bandpass filter to remove double frequency terms produced at the mixer. The signal is now centered at the difference frequency, $w_{IF}-w_{L03}$. It is then resolved into two components in phase quadrature by the two mixers that follow. Finally, the correlation is performed by the integrate and dump circuits shown. As mentioned before, the integrate and dump circuits may be replaced by a suitable low pan filter. In either case, if the ouptut, I and Q, is sampled, it must be sampled synchronously with the data modulation on the signal, or the data modulation must be removed by some means, such as squaring the output.

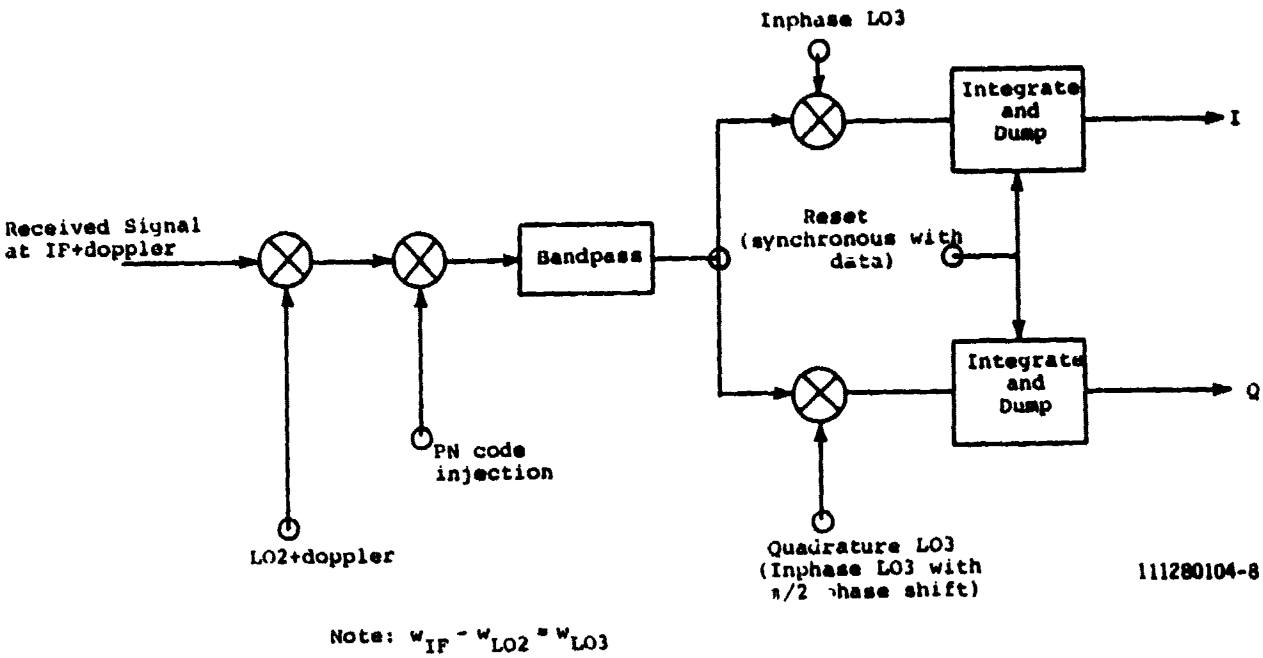


FIGURE 2-7 CODE CORRELATOR AND PHASE DETECTOR

In order to discuss the operation of this stage in more detail, Figure 2-7 has been simplified to the functionally equivalent circuit shown in Figure 2-8. Here, the PN modulation, p(t), is injected first, and the result is translated to zero hertz center frequency in the inphase and quadrature arms. The low pass filters shown with impulse response h(t) perform the correlation as well as remove double frequency terms.

The input signal, s(t), consists of a generic GPS signal and an additive interference term, n(t). The GPS signal is represented as the product of data modulation, d, PN code, P, and a sinusoidial carrier term. In the figure, t represents user time, τ represents all signal delays with respect to user time, and $\emptyset$ represents the corresonding phase delay of the carrier. Both $\emptyset$ and τ will, in general, be time varying, and are related to one another by

$$\tau = \frac{\phi}{2\pi f_c}$$

where f_c is the carrier frequency of the link (L1 or L2).

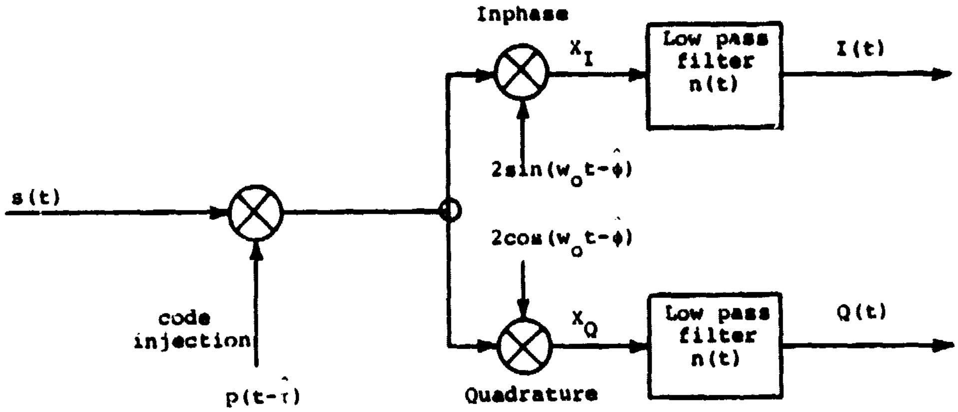
The signal amplitude is A, and average power is $1/2A^2$. The additive interference n(t) is simply represented by its inphase and quadrature components, n_I and n_a. The operating frequency, w_o, can for convenience be taken to be the carrier frequency.

The code and carrier replicas are generated using the receiver's estimates of τ and $\emptyset$, $\hat{\tau}$ and $\hat{\emptyset}$, respectively. The result of the product operations are shown in Figure 2-8, denoted by x_I and x_Q, and are inputs to the low-pass filters. The functions x_I and x_a can be seen to consist of three terms, the first, a deterministic term centered at the difference frequency, zero hertz; the second, a product of the PN sequence and the interference; and the third, a double frequency term, abbreviated f or g. The operation of the low pass filter on the double frequency term essentially eliminates it. We will consider the other two terms separately.

In general, the effect of the low pass filters on x_I and x_Q can be found by convolution with the inpulse response h(t), as shown in the figure. Ignoring the interference terms for the moment, we will examine the effect of the low pass filter on the deterministic portion of the signal. First, we define the error terms $\tilde{\tau}$ and $\tilde{\emptyset}$ as:

$$\tilde{\tau} = \hat{\tau} - \tau$$

$$\tilde{\phi} = \hat{\phi} - \phi$$



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$$S(t) = A\, d(t-\tau)p(t-\tau)\sin(w_o t-\phi) + n(t) \qquad I(t) = n(t) * x_I(t)$$

$$n(t) = n_I(t)\sin w_o t + n_Q(t)\cos w_o t \qquad Q(t) = n(t) * x_Q(t)$$

$$X_I(t) = A\, d(t-\tau)p(t-\tau)p(t-\hat{\tau})\cos(\hat{\phi}-\phi) + p(t-\hat{\tau})n_I(t) + f(2w_o)$$

$$X_Q(t) = A\, d(t-\tau)p(t-\tau)p(t-\hat{\tau})\sin(\hat{\phi}-\phi) + p(t-\hat{\tau})n_Q(t) + g(2w_o)$$

FIGURE 2-8 SIMPLIFIED CORRELATOR/PHASE DETECTOR

where $\tilde{\tau}$ is the error in the local estimate of τ, and $\tilde{\theta}$ is the error in the local estimate of θ. As an example of a low pass filter, we will take h(t) to be:

$$h(t) = \begin{cases} 1/T \text{ for } 0 < t \leq T \\ 0 \text{ otherwise} \end{cases}$$

This type of filter is sometimes called a smoother. Next, we will assume that $\tilde{\tau}$ and $\tilde{\theta}$ are essentially constant over the interval T, that T is less than the duration of one data bit (20 milliseconds), and that T is much greater that the duration of one PN code bit (~ 100 ns for P, 1 μs for C/A). We can then approximate the convolution as:

$$I(t) = Ad(t-\tau)\ \cos\tilde{\theta}\{1/T \int_{t-T}^{t} p(t-\hat{\tau})p(t-\hat{\tau})dt\}$$

$$Q(t) = A\ d(t-\tau)\ \sin\tilde{\theta}\{(1/T \int_{T-T}^{t} p(t-\hat{\tau})p(t-\hat{\tau})dt\}$$

Since the integral approaches a true correlation as T becomes large, the above can be written, still more approximately, as:

$$I(t) \approx A\ d(t-\tau)\ R(\tilde{\tau})\cos\tilde{\theta}$$

$$Q(t) \sim A\ d(t-\tau)\ R(\tilde{\tau})\sin\tilde{\theta}$$

where R is the correlation function of the PN code described previously (Figure 2-1).

Note that if the output of the filter is sampled and held at data transitions, T can be equal to 20 ms without distorting the data modulation. The filter would then be equivalent to an integrate and dump circuit. Also note that as T increases, the bandwidth of the filter decreases. This means that $\tilde{\tau}$ and θ must be more nearly constant for the above relations to hold. That is to say, both the data modulation and the anticipated doppler error $(d\tilde{\theta}/dt)$ must be within the passband of the filter. This also implies that during acquisition, both frequency and delay must be determined in a two-dimensional search process. [CAHN78, p. 5-32].

We turn now to consideration of the interferance term. We will consider two limiting cases--that of narrowband interference, and that of wideband noise. Before proceeding, we will state two results, without proof, concerning the product of stochastic processes. If x and y denote two independent stationary random processes, and z is

their product, then the autocorrelation of z is related to that of x and y by:

$$R_z(\tau) = R_x(t)R_y(T) \quad , \qquad z = xy$$

where

$$R_a(T) = E\{a(t)a(t+\tau)\},$$

the autocorrelation of the process a. Also, since spectral density of a process a is defined as the Fourier transform of R_a, the spectral density of the product can be obtained by convolution:

$$S_z(w) = S_x * S_y = \int S_x(w_1)S_y(w_1-w)\ \frac{dw_1}{2\pi} \quad .$$

[PAPOULIS65, Chapter 10]

Case 1, narrowband interference: A coherent interfering signal at the carrier frequency of the GPS signal can be represented by $n_I = B$, $n_a = o$, where B is a constant. Its autocorrelation is also a constant, B^2, and its power spectral density is $2\pi B^2\delta(w)$. The autocorrelation of p(t) is a given in Figure 2-1, and ignoring the periodicity of the sequence, its spectral density is:

$$S_p(w) = \left\{\frac{\sin(wt_{i/2})}{wt_{i/2}}\right\}^2 \quad .$$

Note that $S_p(o) = t_c$, the duration of one PU code chip.

Therefore, the process $z = p(t-\tau)n_I(t)$ has spectral density:

$$S_z(w) = B^2\left\{\frac{\sin(wt_{c/2})}{wt_{c/2}}\right\}^2 \quad ; \; S_2(0) = B^2t_c$$

and autocorrelation:

$$R_z(\tau) = B^2\phi_p(\tau)$$

The spectral density of the filtered process is given by:

$$S(w) = |H(jw)|^2 s_z(w)$$

where $H(jw)$ is the system function of the lowpass filter (corresponding to $h(t)$). The low pass filter will, in general, have a much narrower bandwidth than the process z, and the spectrum S_z will be essentially flat over the extent of $|H(jw)|^2$. Therefore, one can write:

$$S(w) \approx |H(jw)|^2 B^2 t_c$$

The variance of the output process is then:

$$\sigma^2 \simeq B^2 t_c \int_{-\infty}^{+\infty} |H(jw)|^2 \frac{dw}{d\pi} \; B^2 t_c B_f$$

where B_f is the filter noise bandwidth. B_f is generally on the order of $1/t_d$, where t_d is the data modulation bit time. The ratio T_d/t_c is sometimes referred to as the processing gain of the wideband modulation, for the interference of power proportional to B^2 now has power proportional to $B^2 t_c/T_d$ after the correlation process.

Case 2, wideband interference: If n_I and n_Q are independent, zero mean random processes with constant power spectral density N over the extent of $S_p(w)$, the autocorrelation of n_I and n_Q is $N\delta(\tau)$, and the statistics of the products $p(t)n_I(t)$ and $p(t)n_Q(t)$ are given by

$$R_z(\tau) = \theta(\tau)N_o\delta(\tau) = \theta(o)N_o S(t) = N_o\delta(T)$$

and

$$S_z(w) = N \int_{-\infty}^{+\infty} S_p(w) \frac{dw}{d\pi} = N \quad .$$

The spectral density of the output of the low pass filter is then:

$$S(w) = |H(jw)|^2 N \quad ,$$

and the process has variance:

$$\sigma^2 = N B_f \quad .$$

Note that in the case of wideband noise, the spread spectrum modulation offers no advantage over conventional modulation techniques in terms of noise reduction. The processing gain mentioned above refers only to interference that has a bandwidth narrower than that of the PN code.

It will be assumed here that the environment of the orbital navigator is a benign one in regard to narrow band interference. In other words, no jamming threat is assumed, and radio frequency interferences from other systems aboard the spacecraft (such as L band radar, communications systems, etc.) is nonexistent or intermittent. The noise sources of interest are then wideband noise, such as thermal noise at the antenna, and noise generated in the front end of the receiver. The output of the correlator/phase detector is then:

$$I(t) = A\, d(t-\tau)\, \phi_g(\tilde{\tau})\, \cos(\tilde{\phi}) + n'_I(t)$$

$$Q(t) = A\, d(t-\tau)\, \phi_g(\tilde{\tau})\, \sin(\tilde{\phi}) + n'_Q(t)$$

where n'_I and n'_Q are zero mean random processes with variance $N B_f$. N can be determined from the thermal noise present at the antenna and the noise figure of the receiver.

Signal Tracking

The necessary processing required to detect the GPS signal data modulation and amplitude has been described in the previous section. The correlator/phase detector which performs this operation requires local estimates of the incoming signal's phase and code delay for proper operation. This section will discuss the generation and maintenance of these local estimates, which once obtained, are principal outputs of the receiver--measurements of signal phase and delay.

The technique used to track signal phase or delay is generally a variant of the phase locked loop shown in Figure 2-9. Figure 2-9 is the standard phase locked loop [VITERBI66]. Here the phase (θ) of a sinusoidal input is tracked by generating a replical ($\hat{\theta}$) in phase quadrature. A multiplier is used as a phase error detector. The product

consists of the sum of two sinusoids: one has the phase difference as its argument, the other has the phase sum. This is followed by a linear filter which attenuates the sum term, producing an error signal e. If the phase estimate lags, the phase of the input signal, e, will increase the voltage controlled oscillator's (VCO) frequency, thus reducing the phase error. The loop's VCO thus tracks the phase of the input signal. Figure 2-10 is a linearized phase locked loop (PLL). Here, the sinusoidial nonlinearity is ignored (e− $\hat{\theta}$ is assumed small) and the phase sum term is ignored. Thus, the multiplier is replaced by a difference, and the VCO by an integrator. Notice that signal amplitude directly affects the loop gain of the phase locked loop, and must therefore be maintained constant by an automatic gain control function (AGC) prior to the loop.

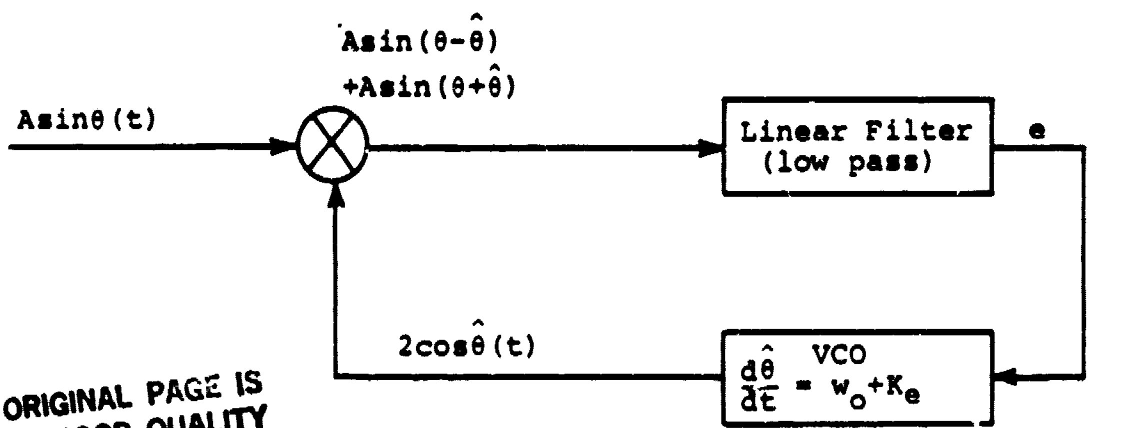

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FIGURE 2-9a PHASE LOCKED LOOP

Tracking loops used to track phase and delay in a GPS receiver differ from the standard PLL primarily in the requirement to eliminate modulation that would prevent operation of the PLL, thus resulting in phase detectors different from the product operation at the standard PLL.

Carrier Tracking

If one examines the outputs of the correlator/phase detector (Figure 2-9), several possibilities for deriving an

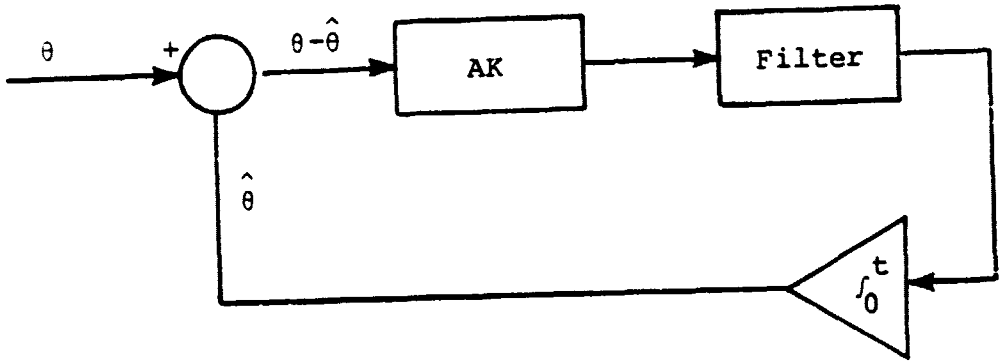

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FIGURE 2-9b LINEARIZED PHASE LOCKED LOOP

estimate of carrier phase error are evident. Some of these are:

a) Square Law:

$$Q^2(t) = \{A\ d(t-\tau)\ R(\hat{\tau})\ \sin\tilde{\phi}\}^2$$

$$= (A\ R(\tilde{\tau}))^2\ 1/2\,(1-\cos 2\tilde{\phi})$$

b) Cosins:

$$I(t)\ Q(t) = \{A\ d(t-\tau)\ R(\tilde{\tau})\ \cos\tilde{\phi}\}\{A\ d(t-\tau)\ R(\tilde{\tau})\ \sin\tilde{\phi}\}$$

$$= (A\ R(\tilde{\tau}))^2\ 1/2\ \sin 2\tilde{\phi}$$

$$\simeq (A\ R(\tilde{\tau}))^2\ \tilde{\phi} \qquad \text{for small } \tilde{\phi}$$

c) Arctangent:

$$Q(t)/I(t) = \tan\tilde{\phi}$$

The most commonly used detector in GPS receivers is the Costas detector. A Costas loop is shown in Figure 2-12, along with a variant of the Costas loop called the Decision Directed Feedback Costas loop.

The noise performance of the Costas loop is degraded from that of the standard PLL, as the noise term driving the loop results from the product of the noise terms in the inphase and quadrature signals.

The Decision Directed Feedback Costas Loop [NATALI69] estimates the data term in the inphase signal, assuming the loop is near phase lock, and removes the data from the quadrature signal by inverting its sign if d(t) is -1. This produces noise performance approaching that of the standard PLL with high signal-to-noise ratios. When the noise power is large enough to create errors in estimating the data bit, however, performance degrades rapidly.

Frequency Tracking

There are two reasons that a GPS receiver would require a carrier frequency tracking function. The first is that it can be an aid during acquisition, not only when carrier phase is unknown, but to some extent, when frequency is unknown. The second is that, under low signal-to-noise ratio conditions, a frequency tracking loop will continue to operate after a phase tracking loop has lost lock. If the code tracking loop is non-coherent (see next section), it can continue to operate as well. A frequency tracking loop cannot estimate the data modulation, however.

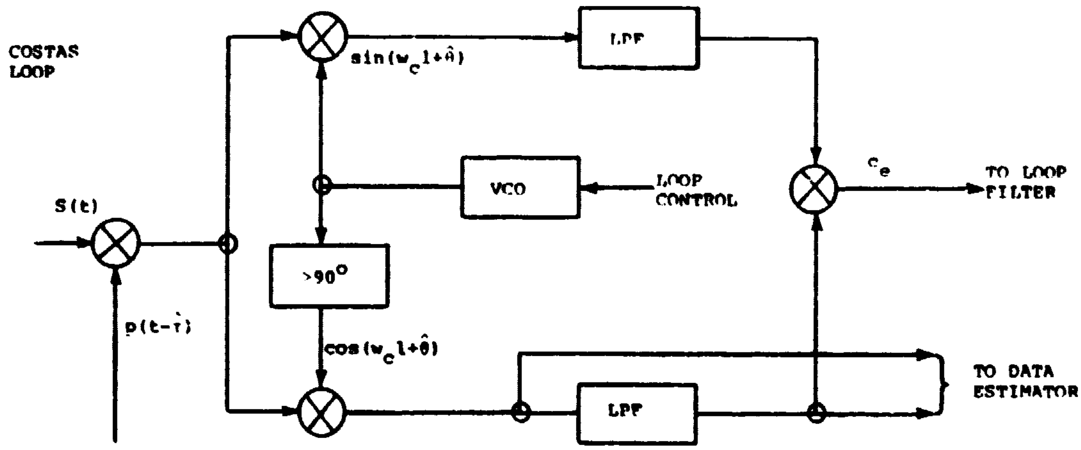
The error discriminator most commonly used for frequency tracking is:

$$I \frac{dQ}{dt} - Q \frac{dI}{dt} = A^2 R^2(\tilde{\tau}) \frac{t}{dt} (\tilde{\phi})$$

or:

$$I(t)Q(t-\Delta) - Q(t)I(t-\Delta) \approx A^2 R^2(\tilde{\tau})\{\tilde{\theta}(t) - \tilde{\theta}(t-\Delta)\}$$

Note that these discriminators, by virtue of performing some type of differentiation, here signularities or spikes at data bit transitions (ignored in the two equations above).



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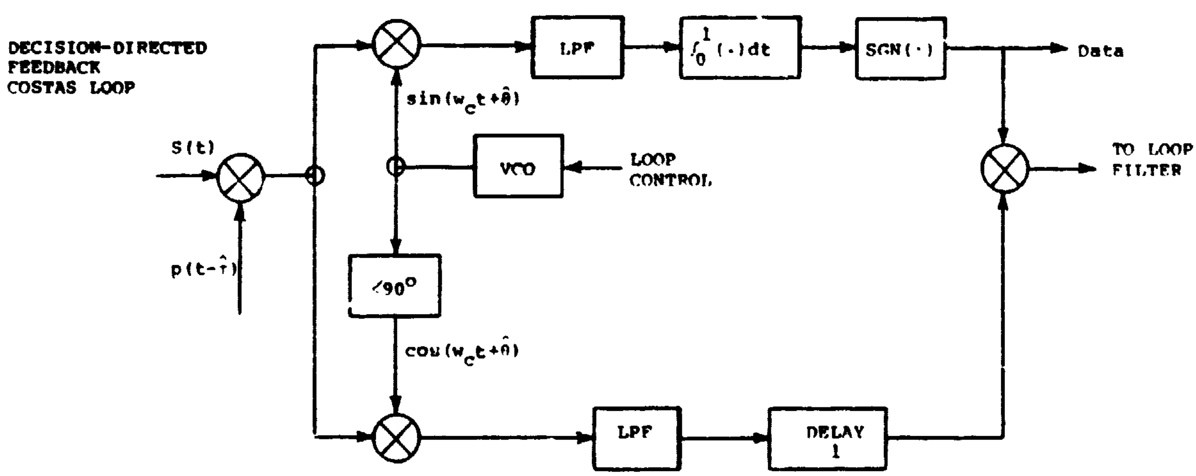

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FIGURE 2-10 COSTAS LOOPS

Code Tracking

There are, in principle, two types of code tracking loop--the coherent code loop and the non-coherent code loop. The coherent code loop assumes in its operation that carrier phase has been accurately determined by the carrier tracking loop (see Figure 2-11). If a correlator is supplied with an accurate carrier phase estimate (including the data modulation), its output would be:

$$I(t) = A\ R(\tilde{\tau}) \quad ; \quad Q(t) = \emptyset \quad .$$

A linear delay error detector can be constructed by correlating the incoming signal with an advanced or "early" code, and a delayed or "late" code, and differencing the result:

$$I_{early} - I_{late} = A(R(\tilde{\tau}+\Delta) - R(\tilde{\tau}-\Delta)$$

A coherent code loop fails when the carrier loop fails to estimate phase. This difficulty can be overcome with the non-coherent code loop as shown in Figure 2-12. Here the output of the early and late correlators are power detected as shown by the squaring and sum operation. The result of the early late difference is, then:

$$\text{Early, Power} - \text{Late Power} = A^2(R^2(\tilde{\tau}+\Delta) - R^2(\tilde{\tau}-\Delta))$$

This detector will be linear only when Δ =1/2 chip.

Note that the non-coherent code tracker does not require the data modulation to be estimated, and in fact will operate satisfactorily whenever the carrier loop's frequency error ($d/dt\ \tilde{\emptyset}$) is sufficiently within the passback of the correlator. GPS receivers typically use non-coherent code loops for this reason. The noise performance of the non-coherent code loop is not as good as that of the coherent loop when the carrier loop is tracking accurately, however. This is due to the squaring operation involved in power detection. Other non-coherent detectors besides the power low detector exist, such as amplitude and peak detectors.

The combined tracking loops (code and carrier) shown in the figures require a total of three correlators. By time multiplexing, or tan dithering, this can be reduced to one correlator for the code loop. Instead of supplying early and late code replicas to two correlators simultaneously, one can provide a single correlator with first an early

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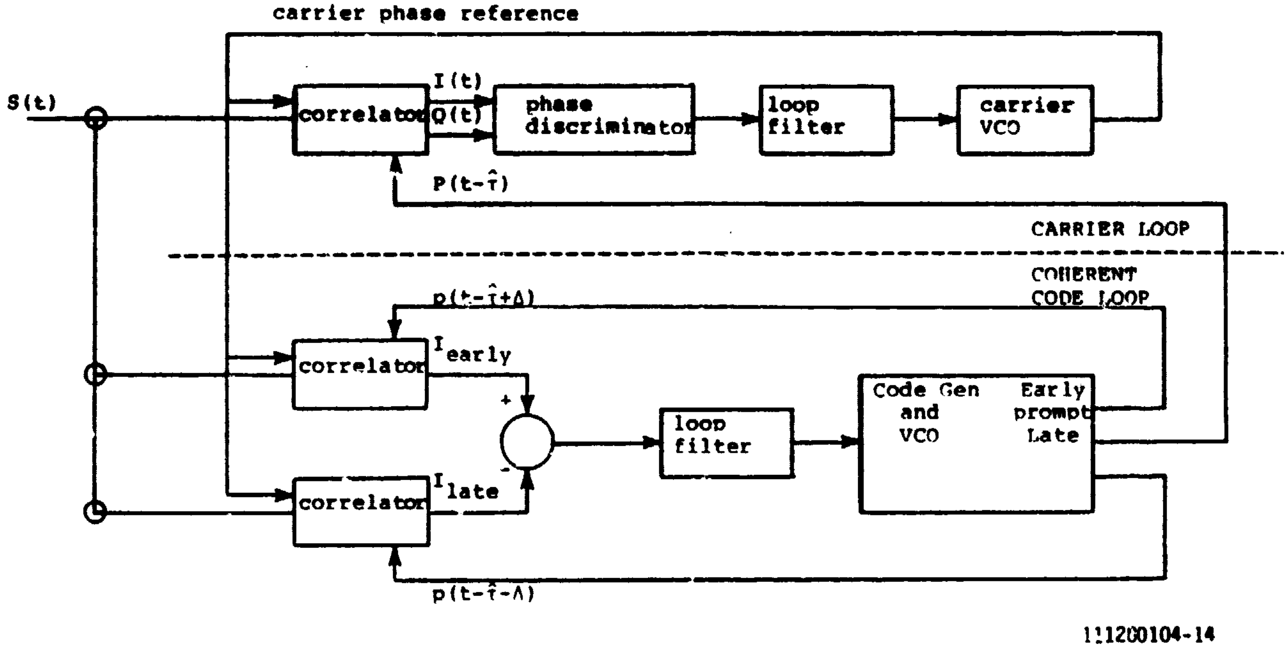

FIGURE 2-11 COHERENT CODE LOOP

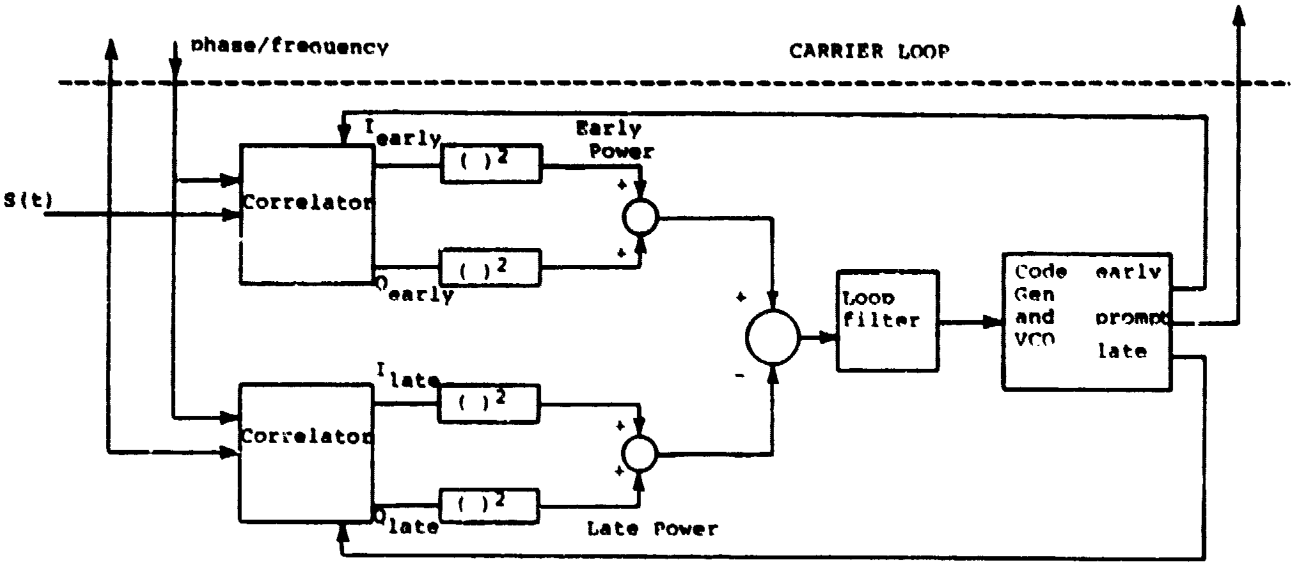

FIGURE 2-12 NON-COHERENT CODE LOOP

code, then a late code. The correlator's output during the early interval is saved, and differenced with its output during the late interval. This will reduce the code tracking loop performance, since total integration time is halved.

Similarly, the requirement for a separate carrier loop correlator can be eliminated. This is done by replacing the inphase input to the carrier tracking loop with the sum $I_{early} + I_{late}$, and replacing the quadrature input with $Q_{early} + Q_{late}$. This results in a performance degradation as well, since the signal-to-noise ratio into the carrier loop is reduced.

Therefore, a single tracking channel can consist of one, two or three correlators, depending upon channel tracking requirements.

Measurement Generation

The previous discussion has described the operation of tracking the GPS signal. By integrating the commands issued by the tracking loops to the VCO's, one obtains the measurements of pseudo-range and pseudo-delta-range. Pseudo-range is defined as the time difference between the time of week indicated by the common system clock and the time of week indicated by the state of the received PN code, at a particular epoch of the system clock. Thus, the pseudo-range for a particular received signal or channel indicates geometric range or time of flight, various atmospheric delays, plus the user clock error (with respect to GPS system time). Pseudo-delta-range is defined as the integrated frequency shift of the user's clock required to maintain phase co-incidence with the carrier of the received signal between two well defined epochs of the system clock.

Data Demodulation

Data demodulation can occur whenever the received signal's carrier is being tracked coherently by the carrier loop. The carrier loop demodulates the data in the following manner. When the loop is tracking, the received signal is either inphase with the local carrier replica, or π radians out of phase (corresponding to $I = A$, and $I = -A$, respectively). These two states represent either a 1 or 0 state of the data or a 0 and 1 state. That is, it is not known whether it is the data that is being demodulated, or the complement of the data. This problem is compounded if the carrier loop "slips a cycle," as this correspondence is likely to have changed, resulting in the loss of not only a single bit, but the complementing of the remainder of the entire message. This ambiguity is common to all biphase, suppressed carrier modulation schemes. The ambiguity is

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resolved by redundant parity encoding of the data. For every 24 bits of data, there are 6 corresponding parity bits. The parity scheme will allow the receiver to detect and correct single bit errors, as well as resolve the biphase ambiguity.

CHAPTER 3

APPLICATIONS SOFTWARE REQUIREMENT

3.1 Applications Processing

This section describes the generic application software functions appropriate to an orbital GPS terminal. The summary presented here forms a basis for the discussion of applications software modularity which appears in a subsequent section of this chapter. These functions can be logically segmented into the following categories:

- Receiver Management and Control
- Measurement Processing
- State Propagation
- Emitter Data Formatting
- I/O Processing

3.2 Receiver Management and Control

As will be discussed in a later section, the primary criterion for placement of a function in this category is that these requirements are enforced by the specific design of the receiver. In the actual design of the applications software, these functions will be the only ones potentially requiring modification when receiver design changes occur. The following functions are discussed below:

- Channel Calibration
- Antenna Management

- Emitter Selection
- Channel Management
- Acquisiton Aiding
- Emitter Data Acquisition

3.2.1 Channel Calibration

This function is required only when mutliple physics channels are present in the receiver, as in the high performance option described in Section 3.3. In this case it is necessary to initially and periodically measure and remove differential biases in the signal tracking loops so that the independent measurements from multiple emitters are time coherent. Once signal tracking is commenced on at least one channel, this function can be conveniently accomplished by causing the remaining channel to track the available emitter. This demonstrates an inherent advantage of the design approach recommended in 3.3--that this function can be standardized across all multichannel designs with the essential characteristics refected in 3.3. This results from the fact that the command and data interface with the receiver is processor-to-processor, and can therefore be fixed from the applications point of view. A specific receiver design can then be characterized by the number of channels, required interval between calibrations, and measured interchannel biases resulting from a coherent read of the tracking loop states.

3.2.2 Antenna Management

This function is required only when multiple antennas are present and the earth-relative orientation of the host vehicle is known. In such a case, it can be determined whether the line-of-sight to a given emitter lies within an antenna gain pattern, and if so, which. If the emitter is visible to an antenna, an appropriate directive is issued to the channel management function; otherwise, a flag is set to inform the emitter selection function that the designated emitter is not currently visible. In the absence of vehicle attitude information, the receiver signal processor is required to listen alternately on the available antennas for the assigned emitter. This function can be partially standardized by fixing the selected antenna command protocol, and providing a configuration-dependent submodule to calculate location of an emitter line-of-sight with respect to each antenna gain pattern.

3.2.3 Emitter Selection

Given the noise and measurement characteristics of the receiver, the maximum reduction in state errors occurs when the navigation processor is fully utilized by the measurement processing function, after all other required functions are allowed for. The rate at which new information can be processed is limited either by the rate at which it can be acquired, or by the rate at which it can be processed. In either case, each measurement or set of measurements represents the expenditure of a scarce resource in the system, and must therefore represent the most effective employment of this resource. This observation defines the function and objective of emitter selection--to assign the available physical (or logical) channels in such a manner that the maximum rate of new information into the navigation processor is maintained.

For an orbiting host vehicle with quiescent body attitude dynamics, the choice of algorithm is less a function of channel multiplicity than the number of available emitters. In view of the recent decision to reduce the constellation density, it seems likely that a simple selection scheme which acquires data from all visible emitters will adequately fulfill the requirements, regardless of set design and channel multiplicity.

3.2.4 Channel Management

This function allocates available physical channels among the current selected emitters, and transmits antenna selection, loop aiding, and data collection commands from the emitter selection, antenna management, acquisition aiding, and data collection functions, respectively. Other than the antenna coverage computation subfunction of antenna management and the emitter data acquisition function, the purpose of this module is to account for all receiver-dependent command and control operations required of the application's processor. It should be emphasized that the modularity characteristics of the application's software inherently depend on the confinement of receiver design-specific software to the minimum number of modules. To the extent that such confinement is achieved, all other application functions remain unaffected by evolutionary changes in signal detection and tracking technology, or become available for incorporation in alternate user equipment.

3.2.5 Acquisition Aiding

Acquisition aiding determines the best estimate of code and frequency offset of the received signal, and determines from the current filter statistics whether direct-P acquisition of the signal should be attempted. This function does not depend on the number, configuration, or method of implementation of the receiver, and therefore can be generically constructed for 1 to n channels.

3.2.6 Emitter Data Acquisition

This function forms the interface between emitter data formatting and the receiver. Its specific operation depends on whether the receiver demodulates and transmits complete data words, performs parity bit stripping and checking, or generates partial subframes of raw data. It is intended to enable a standardized implementation of the emitter data formatting function, operating on not less than 1 complete packed binary data word, which has been parity checked and identified with its proper satellite data block word number.

3.3 Measurement Processing

This segment receives unformatted measurement data from the Receiver Management and Control, and performs all computations to correct the current estimate of the vehicle state and associated statistics. In order to achieve maximum modularity, the included functions are segmented as follows:

- Computation of the Filter Dynamics and State Noise Matrices
- Computation of the Measurement Gradient Matrix
- Computation of the Predicted Value of Pseudo-Range
- Computation of Predicted Value of Integrated Doppler
- Propagation of the Covariance and Update of State and Covariance with Available Measurements

3.3.1 Filter Dynamics, Noise, and Gradient Matrix Computations

The purpose of separately identifying computation functions is to reflect the likelihood that varying mission requirements and host vehicles may require alternate definitions of the estimated state variables. All other functions except these two do not depend in structure or operation on the definition of the filter state, but at most on its dimension, which is specifiable in the data base. Alternate formulations of the estimated state can be co-resident in the application software and invoked as required without impacting the observable prediction, covariance propagation, and state updating functions. It is therefore required that all computations and parameter values specific to the definition of the estimated state be performed and specified in these two functions.

3.3.2 Pseudo-Range Computation

All receiver designs include this function as a basic module, whether or not delta-range measurements are also provided. This function includes computation of the emitter position; corrections for L1/L2 delay, if available; and measurement variances reflecting the state of the tracking loop, C/A or P signal, age of satellite data, and quality of satellite data (that is, ephemeric or almanac).

3.3.3 Integrated Doppler Computations

This function may be inactive (or deleted) in the event the user terminal does not require (or provide) integrated doppler measurements. In the event it becomes active, this function will compute integrals of the satellite and user vehicle position change over the interval, and compute corrections and variances similar to those specified above for the pseudo-range computation.

3.3.4 Covariance Propagation and State Updating

These functions provide for a standard set of covariance propagation and filter update computations, dependent on the dimension of the filter state. The covariance propagation function shall also perform any required extrapolation or interpolation of the state correction to the required state epoch, and incorporation of this correction into the system state. Covariance

propagation as defined here includes incorporation of the process noise matrix.

3.4 State Propagation

This function provides for time propagation of the system state consistent with the accuracy and availability requirements of the solution. In general, the accuracy requirement is determined by the accuracy of the modelling and estimation of non-Toplerian forces and by the information rate sustained by the user terminal during the periods of interest. Since these can vary as a function of orbital shape and dimension and definition of the filter state and perturbing force model, it is unlikely that this function can be made totally generic and standardized. There do seem, however, to be two circumstatnces of sufficiently general application to be fixed as modular options selectable in-flight or pre-launch. The first of these is the relatively low-altitude mission which provides deterministic coverage at all times. In this case, the information rate reduces the rqeuirement for accurate disturbing force modelling, and admits an economical propagation algorithm such as earth-fixed cartesian. In the second case, high-apogee orbits may require more extensive modelling of non-central forces and a correspondingly more robust propagation algorithm. In both cases, the objective is that the state propagation shall not be a contributory source of eror above the unmodelled disturbing forces. [We note parenthetically that the conclusion regarding low-earth orbits may become invalid for some missions, if the current decision for an 18-satellite constellation remains in force. Appendix A shows that periods of reduced coverage will be encountered briefly for all inclinations, even at low altitudes.]

A further requirement on the propagation function concerns availability. Outputs of the system state, or a partition thereof, may be required synchronously for output, or asynchronously for measurement processing and support of other host vehicle requiements. It is therefore proposed that a separate subfunction should be defined which performs all interpolation, and extrapolation functions rqeuired to support I/O and filter operations. Specifically in regard to the latter, this subfunction shall provide a precise value of the system state at each measurement epoch.

3.5 Emitter Data Formatting

Emitter Data Formatting assembles complete frames of satellite data from single coords or partial subframes, unpacks, scales, and converts the resulting frames to engineering units. In the event that accuracy by data corruption is denied, these effects are removed in this function by application of the appropriate key. This function will monitor the availability and age of all ephemerids associated with current emitter, will predict the occurrence of almanac updates and ephemeris page changes, and will command the data acquisition function to collect required emitter data. This function can be made generic to all user set design, and it is a basic module of each.

CHAPTER 4

RECEIVER HARDWARE/SOFTWARE PARTITIONING ANALYSIS

This chapter will examine the question of Hardware/Software partitioning. The generic GPS receiver functional requirements have been described, and a choice must be made as to which of these functions are to be performed by software executed on a mini or microcomputer, and which are to be performed by dedicated signal processing hardware. This partitioning choice will be considered in terms of its effect on the following properties of the resultant system:

System Flexibility: The orbital navigator will be expected to operate in several different environments--different host vehicles, different orbital classes, and different accuracy requirements. While no single receiver would satisfy all mission requirements, it should be possible to reconfigure the system for a specific application.

System Legacy: Many of the component parts or functions of a GPS receiver are not expected to change over the life cycle of a receiver system; yet if care is not taken, many of these functions may have to be rebuilt if a system upgrade is undertaken.

System Development Cost: This will not be a prime consideration here, but development costs must be bounded.

System Modularity: This is not necessarily an end in itself, but generally has a direct effect on flexibility, legacy, and development cost. System modularity implies functional independence and well-defined interfaces between functions.

System power consumption, mass, and volume requirements: Requirements of low power consumption, mass and volume, may in some cases compete with the other properties listed here. In general, a trade-off must be made.

System reliability: This is always a consideration for an orbital system and stringent requirements here will have an adverse effect on other system properties, although system modularity can be used to advantage in improving reliability. It is assumed that the orbital navigator's reliability requirements are consistent with those of a non-flight critical system.

4.1 Partitioning Alternatives

This section will examine a range of partitioning alternatives for the hardware/software boundary. Figure 4-1 is a high level functional breakdown of a generic GPS receiver. The antenna and frequency conversion stages are discussed in the section on "signal acquisition"; the correlator/phase detector is described in the section on signal detection; and the carrier/code tracking and AGC loop filters are described in the section on signal tracking. Measurement generation consists of accumulating carrier VCO commands to form pseudo-delta range, and code generator commands to form pseudo-range, which are then available to the Navigation and Control Function as measurements. Below each section, the range of processing rates required to perform the function is given. The term "processing rate" is meant to indicate the rate at which some repetitive function must be performed, and it does not indicate the "effort" that must be expended at each repetition. For example, the 1 KHz processing rate of the carrier loop would correspond to a handful of integer operations per cycle, while the .1 Hz rate of the navigation filter might correspond to several thousand floating point operations per cycle. The processing rate is directly related to the bandwidth of the signal that a particular function must handle.

Also shown in Figure 4-1 are typical data rates at two interfaces of interest.

The 32K BPS data rate and the 1 KHz processing rate at the correlator/tracking loop interface are maximum rate requirements that occur during signal acquisition. During bit synch, only C/A epochs at 1 ms intervals are known, and there is a 20 ms ambiguity in knowledge of the data bit transition. Loss of information would result if the correlator were to average over a data bit transition, hence, the 1 KHz rate. After this ambiguity is resolved, the processing rate can be reduced as allowed by unknown signal dynamics (the bandwidth must be higher than the frequency errors expected during tracking). The lower limit is 50 Hz, the data modulation rate. The data modulation must be removed in the carrier loop.

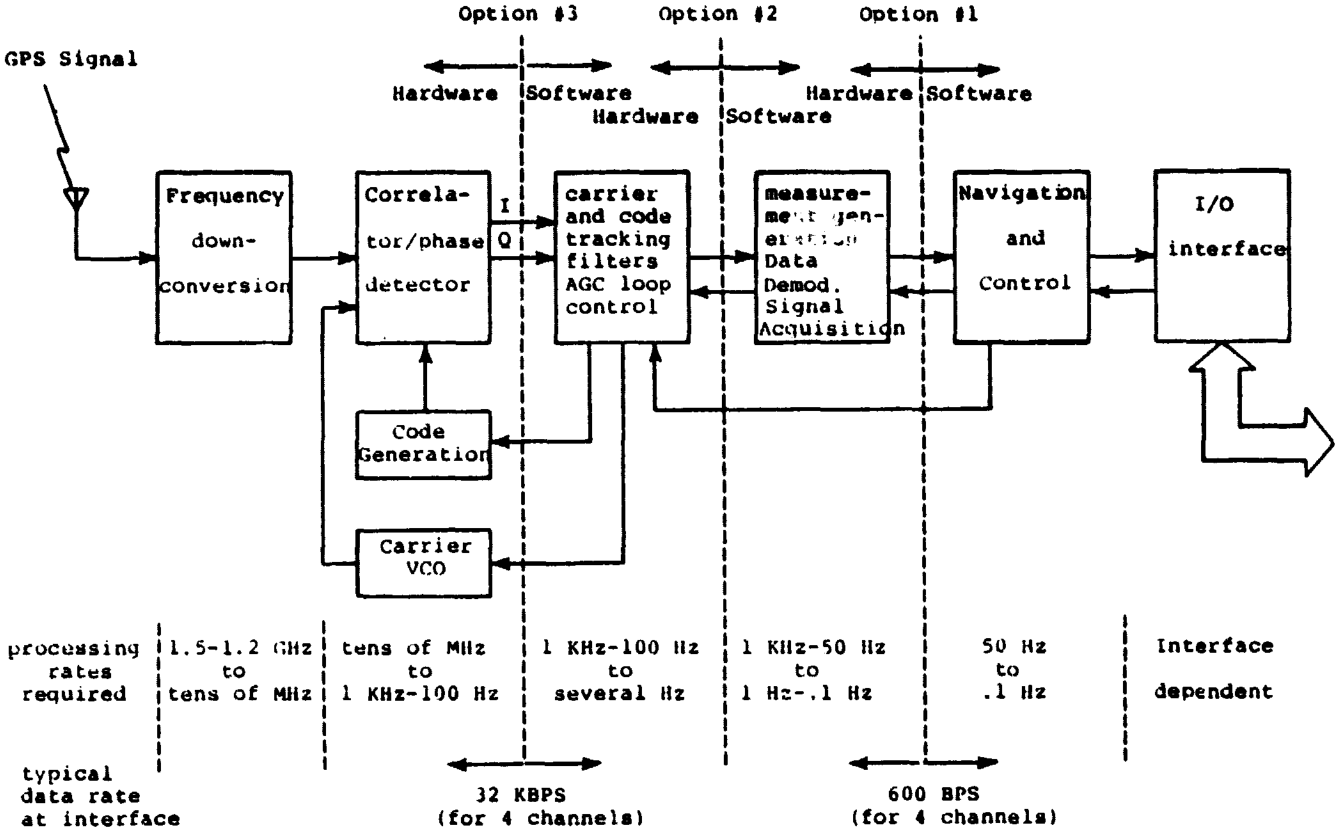

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FIGURE 4-1 PARTITIONING ALTERNATIVES

At the top of Figure 4-1, three partitioning options are shown. Options 3 and 1 indicate upper and lower bounds. Option 3 corresponds roughly to the limit of current microcomputer technology--such machines currently have basic instruction cycle times on the order ofd five microseconds (including instruction and operand fetch) and 16-bit hardware multiply times of 25 to 50 microseconds. With a machine of this type (LSI-11, TI-9900), it is possible to collect I and Q samples at a 1 KHz rate, preprocess them (to remove the data ambiguity) and perform carrier loop processing at a 250 Hz rate. It is unlikely that the correlator/phase detector processing at 2 to 20 MHz would be possible with a low power micro processor in the near future.

At the other end of the spectrum, it is feasible to perform all signal processing with dedicated digital and analog signal processing hardware until one has to compute a GPS navigation solution. This partitioning scheme is represented by option 1 in Figure 4-1.

Option 2 is intended to represent a compromise between options 1 and 3.

Option 3 was used by the Magnavox X set receiver. It made use of a mini-computer for four simultaneous receiver channels, and a second mini-computer for navigation, control and the I/O interface.

The Missile Accuracy Evaluator (MAE) being developed by Charles Stark Draper Laboratories uses this partitioning scheme in conjunction with a federated, micro-processor per signal channel architecture.

"Lost cost" receivers, such as the Y and Z sets, typically use a partitioning scheme somewhere between options 1 and 2.

4.2 Trade-Off Analysis

Figure 4-2 shows a breakdown of system trade-offs against selection of partitioning optin 3 (software baseband signal processing) versus options 1 and 2 (hardware signal processing).

The principal point here is that the GPS signal structure imposes a large number of functional requirements for successful signal acquisition and tracking. (That is, carrier phase tracking, carrier frequency tracking, code delay tracking, variable loop bandwidth, coherent and/or non-coherent AGC, data demodulation, all coordinated with signal acquisition logic.) The decision to perform these functions in software results in fewer hardware requirements (that is, in a microprocessor, memory and interface) traded for more complex software requirements.

Trade-Off	Option #3 Software Signal Proc.	Option #1, 2 Hardware Signal Proc.
Legacy and Flexibility	Can be very good. Can be very flexible with regard to changes in tracking philosophy. Gracefully accommodates complex multi-mode or adaptive bandwidth tracking and acquisition designs.	Tracking philosophy is fixed in hardware, and changes may require board redesign. Bandwidth adaptation requires additional hardware and switching logic.
Modularity	Very high. Tracking channel and Navigation processors may be identical Boards, or even the same µP.	Can be good, but the number of functionally independent hardware modules in the system is much higher than option 3.
Power Consumption Mass, Volume	Power consumption is generally higher than options 1, 2 (a µP module + memory -- 5-20 watts). Mass and Volume requirements may be lower due to reduced board and chip counts.	Power consumption can be very low, making use of micro-power analog circuitry, CMOS logic, etc. (~several watts per tracking channel).
Reliability	There are frew suitable, commercially available space qualified µP's, This situation is expected to improve in the near future.	Space qualified hardware is available--however, high board and chip counts have adverse effect on reliability.
Development Cost	Hardware development traded for software development. In principle, software development cost should be lower, but industry experience seems to indicate a degree of unpredictability here.	Due to large number of functional requirements imposed by the GPS navigation signal, hardware signal processing development may be somewhat high in comparison with other hardware development costs.

FIGURE 4-2 HARDWARE/SOFTWARE TRADE-OFFS

4.3 Recommended Approach

Due to advantages in system modularity, legacy and flexibility implied by software signal processing, that is the recommended approach. Figure 4-3 shows the generic architecture of such a system. Figure 4-4 gives two potential receiver configurations possible with such a scheme. Figure 4-4a is a high performance, multi-channel set, incorporating a micro-processor per tracking channel design. It has L1/L2 tracking capability, as well as C/A and P code tracking capability, and incorporates multiple independent correlators per channel. Figure 4-4b is a low performance, single channel set. It makes use of the L1 C/A link only, and uses a single time-multiplexed correlator. The functions of signal tracking and navigation are performed by one microprocessor. The increased computational burden that this imposes on the microprocessor implies sequential performance of measurements (signal tracking) and navigation cycles (Kalman filter updates) at a much reduced rate.

These configurations represent extreme limits of hardware versus performance requirements. There are many intermediate configurations possible.

The hardware and software commonality implicit in Figure 4-4 should be stressed. The micro-processors shown in the figures could all be of identical design, differing only in memory and interface requirements. Navigation and tracking software could be largely identical as well, with only the control and interface logic being different for each configuration.

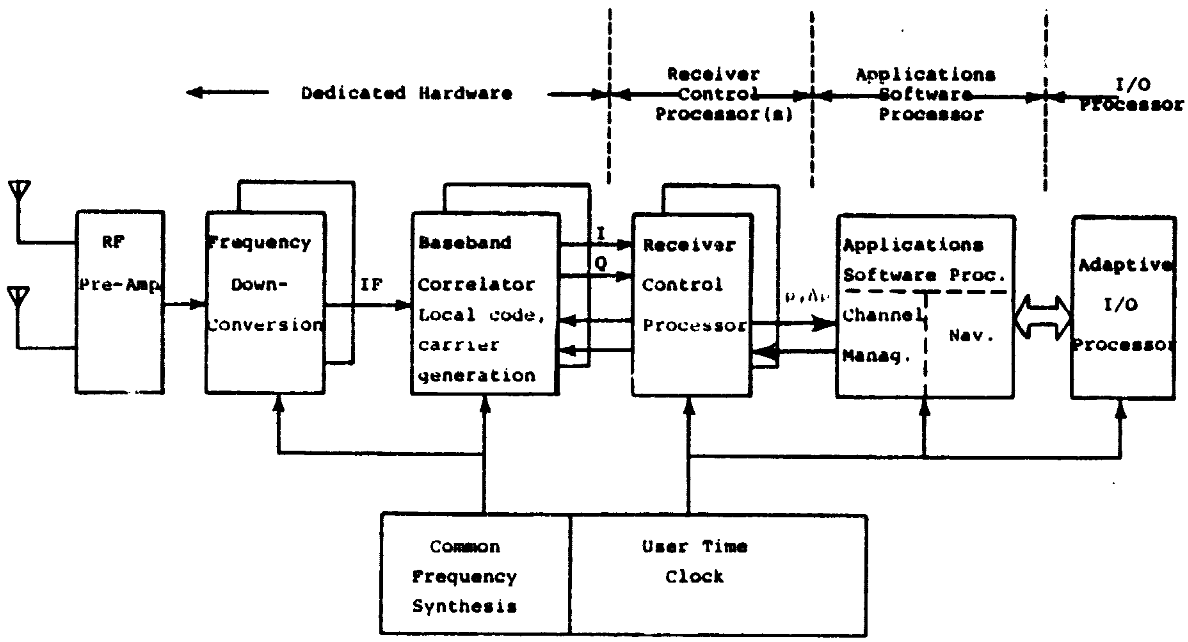

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FIGURE 4-3 GPS ORBITAL USER TERMINAL FUNCTIONAL BREAKDOWN

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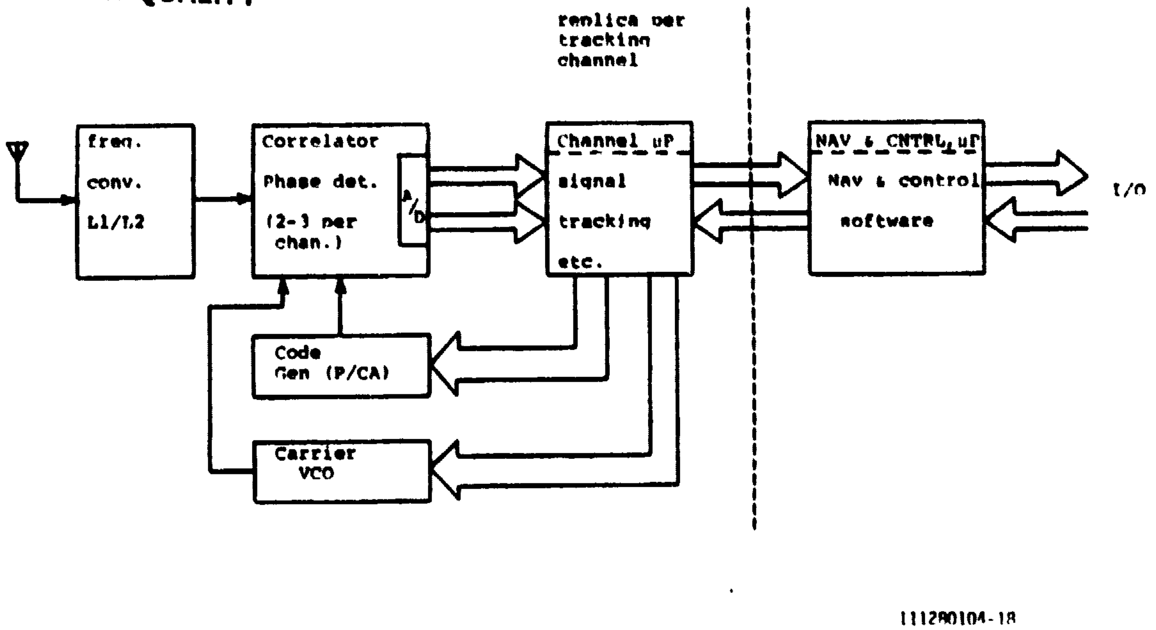

FIGURE 4-4a MULTIPLE CHANNEL HIGH PERFORMANCE SET

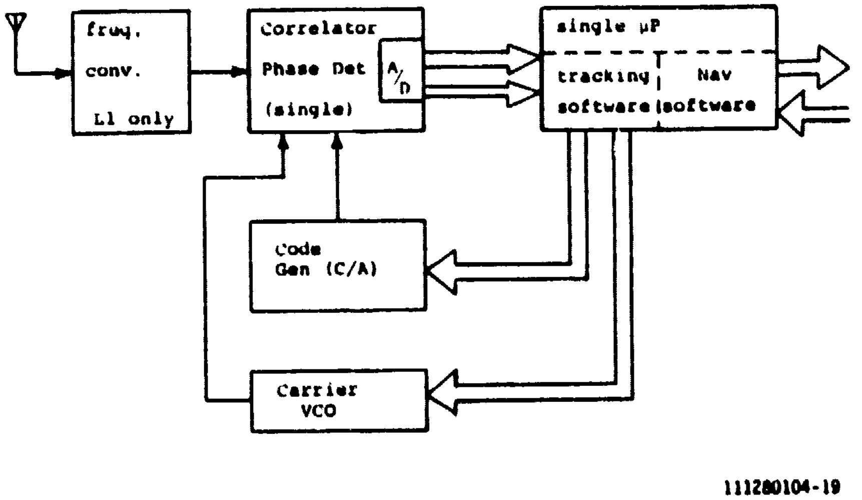

FIGURE 4-4b SINGLE CHANNEL LOW PERFORMANCE SET

FIGURE 4-4 MICRO-PROCESSOR ORIENTED RECEIVER CONFIGURATIONS

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